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CHAPTER 4

EXTERNAL INTERFERENCE

4.1 INTRODUCTION

Microwave telecommunications radio relay systems all over the world have frequencies greater than 1 GHz. Many factors must be considered in the proper choice and utilization of the various frequency bands. Each band has advantages as well as limitations.

An appropriate segregation of various bands for various users is useful for optimum frequency utilization. It is assumed that various bands will be set aside for similar telecommunications users. This is necessary to allow simplifying assumptions to be made regarding transmit powers, interfering spectrums, and receiver susceptibility. It is difficult to coordinate the use of low- and high-power terrestrial microwave equipment in the same band and geographic region.

Once the telecommunications bands have been decided upon, a choice must be made as to the use of that band for digital or analog traffic and whether that traffic will be low- (less than 600 channels) or high-density transmission. In general, low- and high-capacity systems should not be allowed into the same band. One system will soon use up frequencies that are needed by the other. As a matter of principle, it must be assumed that any given band will eventually become fully expanded. Band splitting and avoiding certain frequencies are undesirable. If two systems operate in the same band and same geographical area, interference is minimized by having them on the same plan. Tables 4-1 through 4-6 list typical microwave terrestrial and satellite transmission bands.

A microwave network that is very dense, with many branching points and crossings, can take advantage of a low- to medium-density frequency plan. A microwave route with few branching points is generally most economical using a high-density plan. The guiding principle of channel assignment planning is to use as few frequencies within a band as is necessary. Those frequencies are then repeated as often as possible to retain the maximum growth potential. In this regard, designating two groups of frequencies within a band for use in a high/low pattern is called a 2-frequency plan. This is because only two sets of frequencies are used anywhere on the route. If another two sets of frequencies are used (eg, main and interleaved together), the frequency grouping is called a 4-frequency plan.

Table 4-1. ITU terrestrial band allocations (1 to 25 GHz).

Center Frequency (GHz)	Band (GHz)	Region 1 (Europe, Africa, USSR, Turkey, Mongolia)	Region 2 (North America, South America)	Region 3 (Far East, Australia)
1.375	1.350 - 1.400	*	-	-
1.481	1.427 - 1.535	*	*	*
1.67525	1.6605 - 1.690	*	*	*
1.695	1.690 - 1.700	*	-	-
2.195	1.700 - 2.690	*	*	*
3.350	3.300 - 3.400	-	*	-
3.800	3.400 - 4.200	*	*	-
4.700	4.400 - 5.000	*	*	*
7.175	5.850 - 8.500	*	*	*
10.225	10.000 - 10.450	*	-	*
10.590	10.500 - 10.680	*	*	*
11.600	10.700 - 12.500	*	*	*
12.625	12.500 - 12.750	-	*	*
13.000	12.750 - 13.250	*	*	*
14.350	14.300 - 14.400	*	-	*
14.600	14.400 - 14.800	*	*	*
18.700	17.700 - 19.700	*	*	*
22.400	21.200 - 23.600	*	*	*

SOURCE: ITU Radio Regulations, 1982, Volume I, Part A, Chapter III, Article 8, Sections I and IV.

Table 4-2. CCIR terrestrial band allocations.

Center Frequency (GHz)	Band (GHz)	Use			CCIR Rec.
		Analog Telephony	Digital	Analog Television	
1.800	1.700 - 1.900	*	*	-	283-4
1.908	1.708 - 2.108	*	-	*	382-4
1.932	1.732 - 2.132	*	-	*	382-4
2.0865	1.8865 - 2.2865	*	-	*	382-4
2.000	1.900 - 2.100	*	*	-	283-4
2.101	1.901 - 2.301	*	-	*	382-4
2.200	2.100 - 2.300	*	*	-	283-4
2.600	2.500 - 2.700	*	*	-	283-4

See source at end of table.

Table 4-2. CCIR terrestrial band allocations (cont).

Center Frequency (GHz)	Band (GHz)	Use			CCIR Rec.
		Analog Telephony	Digital	Analog Television	
3.800	3.400 - 4.200	-	*	-	635
3.950	3.700 - 4.200	*	*	*	382-4
4.0035	3.8035 - 4.2035	*	*	*	382-4
6.175	5.925 - 6.425	*	*	-	383-3
6.770	6.430 - 7.110	*	*	-	384-4
7.275	7.125 - 7.425	*	-	-	385-3
7.400	7.250 - 7.550	*	-	-	386-3
7.575	7.425 - 7.725	*	-	-	385-3
7.700	7.550 - 7.850	*	-	-	385-3
8.000	7.725 - 8.275	*	-	-	386-3
8.350	8.200 - 8.500	*	-	-	386-3
11.200	10.700 - 11.700	*	*	*	387-4
13.000	12.750 - 13.250	*	*	*	497-2
14.875	14.400 - 15.350	-	*	-	636
14.925	14.500 - 15.350	-	*	-	636
18.700	17.700 - 19.700	-	*	-	595
22.400	21.200 - 23.600	*	*	*	637

SOURCE: CCIR Recommendations (Green Books), Volume IX-1, 1982.

Table 4-3. USA terrestrial band allocations (1 to 25 GHz).

Center Frequency (GHz)	Band (GHz)	Use				Notes
		Govt.	Common Carrier	Private Fixed	Television Relay	
1.740	1.710 - 1.770	*	-	-	-	1
1.920	1.850 - 1.990	-	-	*	-	6
2.145	2.110 - 2.180	-	*	-	-	2
2.165	2.130 - 2.200	-	-	*	-	6
2.245	1.990 - 2.500	-	-	-	*	3
2.245	2.200 - 2.290	*	-	-	-	1
2.475	2.450 - 2.500	-	-	*	-	6
2.595	2.500 - 2.690	-	-	-	*	6
3.950	3.700 - 4.200	-	*	-	-	2

See source and notes at end of table.

Table 4-3. USA terrestrial band allocations (1 to 25 GHz) (cont).

Center Frequency (GHz)	Band (GHz)	Use				Notes
		Govt.	Common Carrier	Private Fixed	Television Relay	
4.695	4.400 - 4.990	*	-	-	-	1
6.175	5.925 - 6.425	-	*	-	-	2
6.700	6.525 - 6.875	-	-	*	-	6/9
7.000	6.875 - 7.125	-	-	-	*	3
7.1875	7.125 - 7.250	*	-	-	-	1
7.6375	7.300 - 7.975	*	-	-	-	1
8.100	8.025 - 8.175	*	-	-	-	1
10.615	10.550 - 10.680	-	*	*	-	2/6
11.200	10.700 - 11.700	-	*	-	-	2
12.450	12.200 - 12.700	-	-	*	-	7
12.925	12.700 - 13.150	-	-	*	-	9
12.950	12.700 - 13.200	-	-	-	*	4
12.975	12.700 - 13.250	-	-	-	*	3
13.225	13.200 - 13.250	-	-	*	-	5
13.225	13.200 - 13.250	-	*	-	-	2
14.825	14.500 - 15.350	*	-	-	-	1
18.700	17.700 - 19.700	-	*	*	*	2/3/4/6
21.600	21.200 - 22.000	-	*	-	-	2
22.400	21.200 - 23.600	*	-	-	-	1
22.500	21.800 - 23.200	-	-	*	-	8
22.700	21.800 - 23.600	-	-	*	-	5
22.800	22.000 - 23.600	-	*	*	-	2/5

SOURCE: Code of Federal Regulations, Title 47 (Telecommunication), Chapter I (Federal Communications Commission), Parts as noted below (as ammended through October 1986).

NOTE 1:	Part 2.106	NOTE 4:	Part 78.18	NOTE 7:	Part 94.90
NOTE 2:	Part 21.701	NOTE 5:	Part 94.61	NOTE 8:	Part 94.91
NOTE 3:	Part 74.602	NOTE 6:	Part 94.65	NOTE 9:	Part 94.93

Table 4-4. ITU satellite band allocations (1 to 25 GHz).

Center Frequency (GHz)	Band (GHz)	Use	
		Up-link	Down-link
1.428	1.427 - 1.429	*	-
1.530	1.525 - 1.535	-	*
2.595	2.500 - 2.690	*	*
3.800	3.400 - 4.200	-	*
4.650	4.500 - 4.800	-	*
6.400	5.725 - 7.075	*	-
7.500	7.250 - 7.750	-	*
8.150	7.900 - 8.400	*	-
11.725	10.700 - 12.750	*	*
13.000	12.750 - 13.250	*	-
14.400	14.000 - 14.800	*	-
17.500	17.300 - 17.700	*	-
17.900	17.700 - 18.100	*	*
19.650	18.100 - 21.200	-	*
22.750	22.500 - 23.000	*	*

SOURCE: ITU Radio Regulations, 1982, Volume I, Part A, Chapter III, Article 8, Section IV; and Part B, Chapter VIII, Article 27, Nos. 2509, 2510, and 2511

COORDINATION: In accordance with ITU Radio Regulations, Part A, Chapter IV, Article 11 (RR11), Sections III, IV, and V; Article 12 (RR12), Subsection IIE and Part B, Chapter VIII; Article 27 (RR27); Article 28 (RR28), and Appendix 28 (AP28); Resolution 703 (RES703); and Recommendation 708 (REC708).

Table 4-5. CCIR satellite band allocations .

Center Frequency (GHz)	Band (GHz)	Use	
		Up-link	Down-link
2.595	2.500 - 2.690	*	*
3.650	3.400 - 3.900	-	*
3.950	3.700 - 4.200	-	*
5.975	5.725 - 6.225	*	-
6.175	5.925 - 6.425	*	-
7.275	7.250 - 7.300	*	*
8.000	7.975 - 8.025	*	*
11.075	10.950 - 11.200	-	*
11.550	11.300 - 11.800	-	*
11.575	11.450 - 11.700	-	*
11.975	11.700 - 12.250	-	*
14.250	14.000 - 14.500	*	-
17.400	17.300 - 17.800	*	-

SOURCE: CCIR Recommendations (Green Books), Volume IV/IX-2, 1986

COORDINATION: In accordance with CCIR Green Books, Volume IV/IX-2, Recommendations 355-3, 356-4, 357-3, 358-3, 406-5, and 558-2.

Table 4-6. USA satellite band allocations (1 to 25 GHz).

Center Frequency (GHz)	Band (GHz)	Use	
		Up-link	Down-link
2.5175	2.500 - 2.535	*	*
2.6725	2.655 - 2.690	*	-
3.950	3.700 - 4.200	-	*
4.550	4.400 - 4.700	*	-
6.175	5.925 - 6.425	*	-
6.875	6.625 - 7.125	-	*
7.500	7.250 - 7.750	-	*
8.150	7.900 - 8.400	*	-
10.075	10.950 - 11.200	-	*
11.575	11.450 - 11.700	-	*
11.950	11.700 - 12.200	-	*
12.450	12.200 - 12.700	-	*
13.625	13.250 - 14.000	*	*
14.250	14.000 - 14.500	*	-
17.550	17.300 - 17.800	*	-
18.700	18.360 - 19.040	-	*
18.950	17.700 - 20.200	-	*
19.950	19.700 - 20.200	-	*
20.700	20.200 - 21.200	*	*

SOURCE: Code of Federal Regulations, Title 47 (Telecommunication), Chapter I (Federal Communications Commission), Parts 2.106, 21.701, 25.202, 25.204, 25.252 (table 1), 94.61, 94.63, and 94.77 (as amended through October, 1986).

COORDINATION: In accordance with the above source documents.

It is important that only one frequency plan be established for a particular band. Multiple plans within a country make frequency coordination difficult and force frequency utilization to be inefficient. Use of a CCIR or FCC type plan is strongly recommended. These channel arrangements have been very carefully developed to avoid many internal sources of interference. In particular, the channel spacings have been developed to place the most significant intermodulation products exactly on or exactly between channel center frequencies. Receiver image responses have also been considered. Although care must be exercised in using any plan (especially when using a single antenna for both transmission and reception), use of nonstandard plans typically cause serious intra- and intersystem problems. Standard plans are given in the appendix of this text.

The choice of a band is based on many factors. Path attenuation between isotropic radiators increases with frequency, so at first glance it would appear that higher frequencies have a disadvantage. However, for a given size of antenna, gain increases with frequency. When antenna gain at both ends of the path is considered, total path loss actually decreases as frequency is increased. This tends to be compensated for by additional transmission line loss, lower transmit power, and increased receiver noise figure available at high frequencies. In general, path loss is not a factor in choice of band. Frequency of operation is significant in climates where intense rainfall is prevalent. Use of frequencies greater than approximately 10 GHz in these climates should be limited to short paths.

Terrain and site accessibility may dictate use of passive reflectors (flat billboards). A passive reflector is only effective if its linear dimensions are large compared to the radio wavelength. Due to size limitations, use of passive reflectors at frequencies lower than 4 GHz is seldom practical. Consideration must be given to beam width of the passive reflector radiation pattern. Large reflectors operated at high frequencies have narrow beam width.

A narrow beam width places significant mechanical stability requirements on the support structure to maintain a stable received signal. Narrow beams are also more susceptible to anomalous atmospheric propagation effects. At frequencies above about 10 GHz, manufacturing and installation tolerance limitations sometimes force the use of interference grating (window blind) reflectors (beam benders) rather than flat (billboard) reflectors. Passive reflectors generally have much worse sidelobe radiation patterns than parabolic or horn antennas. As will be shown in chapter 9, sidelobe performance in a given plane can be improved by rotating the rectangular reflector. If the sites are above or below and to the side of the passive reflector, the polarization of the reflected signal will be rotated. The orthogonality of cross-polarized signals will be maintained. Both are merely rotated clockwise or counterclockwise.

The transmit or receive antenna feed horn can be rotated to reachieve optimum cross-polarization on the path using the passive. Generally, this cannot be done on other paths in the same area, since this can significantly increase interference to and from other sites due to loss of polarization discrimination between the paths. The use of passive reflectors should be avoided where efficient frequency utilization is important.

External interference can become a significant factor in band selection. Air surveillance or weather radar installations tend to have radiation that spills into adjacent frequency bands. Also, bands that are harmonically related can have interference. A common offender is radar operating around 1.3 GHz. The third harmonic of this radar can cause interference in the 4-GHz band. Many radars have spurious outputs. Waveguide is a natural high-pass filter; therefore, there is seldom significant interference below a radar operating frequency. In no case should a microwave installation be built near or in line with a radar. If a microwave installation must terminate at a radar station, the 2-GHz band is least likely to be affected by radars using typical frequencies.

Many tropospheric scatter installations have very high power transmitters. Where a microwave route intersects a tropospheric scatter link, use of microwave radios near fundamental or second and third harmonic tropospheric frequencies should be avoided.

The satellite communications service shares many terrestrial microwave frequency bands. The most common interference cases occur at 4, 6, or 8 GHz. Interference coordination at 8 GHz is usually accomplished by avoiding the most popular 8-GHz satellite frequencies. Utilization of this band is relatively small. Transmission from the satellite to the earth occurs in the 4-GHz band. The satellite is power limited and the signal received at the earth station is extremely small. For that reason, radio relay stations transmitting in the direction of an earth station can cause considerable interference.

Typically, in the 4-GHz band there is a need to coordinate terrestrial microwave transmit powers within 500 km over land (800 km over water) from an earth station. The precise distance depends on many factors, including the type of terrain, the horizon angle at the earth station, the satellite position, earth station antenna pattern, and the particular coordination method (eg, FCC or CCIR). This coordination obligation does not mean that it is impossible to use the 4-GHz band within that range; it just means that for every occasion that the 4 GHz is used for microwave terrestrial transmission, the calculations have to be done to see whether interference to the earth stations exceeds established values.

Earth stations transmit to the satellite in the 6-GHz band with powers as high as 1 kW, which can cause interference to terrestrial microwave receivers over substantial distances if the earth station transmitter illuminates a common volume of a terrestrial route. Since the interference mechanism is a phenomenon similar to tropospheric scatter propagation, the required coordination distance is typically only about one-third that of the 4-GHz band. Another consideration of

6 GHz is that terrestrial microwave radio systems can cause interference to satellites in orbit. This is the case whenever a radio link transmits in a direction that intersects the geostationary orbit. For low-angle terrestrial paths, consideration must be given to antenna elevation and atmospheric refractions.

Coordination with satellite stations is accomplished in accordance with FCC Rules and Regulations, Parts 21.701, 25.204, 25.252, 94.61, 94.63, and 94.77; CCIR Green Books, Volume IV-1, Recommendation 465-1 and Volume IV-2/IX-2, Recommendations 355-3, 356-4, 357-3, 358-3, 406-5, and 558-1; and ITU Radio Regulations Part A, Chapter IV, Article II (RR11), Sections III, IV, and V, Article 12 (RR12), Subsection IIE, Part B, Chapter VIII, Article 27 (RR27), Article 28 (RR28), Appendix 28 (AP28), Resolution 703 (RES703), and Recommendation 708 (REC708). General performance characteristics of earth stations and satellites can be found in the above documents and Volume 1 of the INTELSAT Satellite System Operations Guide.

Long-distance terrestrial microwave transmission typically uses the frequency bands between 1 and 10 GHz. With regard to propagation considerations, the 2-GHz band is quite reliable and rain attenuation is insignificant. Due to the relatively large wavelength of the propagating radio signal, signals in this band are less prone to atmospheric multipath, earth bulge blockage, or ducting. Antenna patterns are relatively broad, making antenna alignment and tower rigidity requirements rather lax. On the other hand, the poor antenna discrimination patterns and small front-to-back ratios make use of the band difficult for high-capacity systems. Interference from spur paths and other users is typically a problem on main routes.

Although the high-density 2-GHz band plans have six duplex frequencies, poor antenna front-to-back ratios often force the use of three frequency pairs on one path and three other pairs on the next path to achieve adequate interference isolation (eg, use of a 4-frequency plan rather than the more desirable 2-frequency plan). If the 2-GHz band must be used for high-density transmission, the 1.7- to 1.9-GHz band is usually chosen for low-capacity routes and 1.9 to 2.3 GHz is chosen for high-capacity systems.

The two bands meet at 1.9 GHz. As with the upper and lower 6-GHz bands, at sites where both bands are used, care should be taken to ensure that transmission or reception occurs at the higher portion of the lower band and the lower portion of the higher band. Generally, the 2-GHz band is best suited for low- to medium-density systems.

The 4-, lower 6-, upper 6-, 7-, and 8-GHz bands are quite popular for medium- to high-capacity systems. The 4- and 6-GHz bands are most popular. Worldwide, the upper 6-GHz band is the band used most commonly for 2700-channel FM transmission. Since it is not near any satellite frequency, it is also popular for use at or near satellite earth station terminals.

To achieve the greatest number of wide-band radio channels in the smallest geographical area, all systems should have common transmit frequencies. This is a result of the observation that for an FM system, the lowest interference noise occurs when the interfering signal has exactly the same frequency as the desired signal or a greatly different frequency. Historically, the first approach to a preassigned frequency plan placed transmit and receive channel frequencies for a given path on adjacent frequencies (as frequencies increased, channel use alternated between transmit and receive). It was soon observed that if adjacent channels were placed on alternating orthogonal polarizations (vertical or horizontal), channel capacity could be doubled by putting two transmit or two receive channels where there had been one. Even greater channel spacing efficiency was achieved by grouping all transmit channels at one end of the frequency band, all receive channels at the other end, and providing a band of unused frequencies (guard band) between them. This had the added benefits of relaxing receiver filtering requirements and allowing the use of a single antenna for transmission and reception for moderate numbers of RF channels. Alternating polarization (horizontal followed by vertical) of adjacent frequency channels maximized adjacent channel rejection.

At sites with several converging paths or where economic constraints require the use of lower performance antennas, some paths can be placed on frequencies offset from the main frequency plan by one-half a channel bandwidth (interleaved frequency plan), taking advantage of the reduced interference of two signals with large frequency offset. Although frequency interleaving can be used to solve specific problems, it usually makes frequency coordination with other users more difficult and generally complicates system expansion.

Efficiency of frequency use is improved by antennas with very low front-to-back coupling (eg, horn or shrouded parabolic reflector antennas). Passive reflectors generally have radiation patterns significantly worse than horn or shrouded parabolic reflector antennas. Also, when the passive is above or below either site, the reflected signal experiences polarization rotation. These factors make use of passive reflectors undesirable where maximum frequency reuse is desired.

Using high-performance antennas, it is possible to transmit two directions with the same sets of frequencies. The polarization of one path is often reversed on the next path to minimize path-to-path interference if foreground reflection or ground clutter backscatter (transmitted signals reflected by terrain near the transmitter) are potential problems. A typical 2-freqency microwave system frequency pattern repeats itself every two hops. Interference over three hops will not normally occur; however, abnormal propagation conditions can cause such interference. If this condition is probable, the polarization is changed every two hops (two hops with horizontal polarization followed by two hops of vertical polarization). Additional discrimination (large antennas or zig-zag paths) may be necessary if the microwave route is approximately in a straight line.

For frequency bands where multiline operation with a common antenna is the norm, the guard band between transmit and receive subbands is sometimes chosen as an odd-half multiple of the channel bandwidth (typically 1/2 or 3/2). This causes most multiple transmitter intermodulation products to fall halfway between the receiver channel frequencies, resulting in minimal demodulated baseband interference.

Occasionally, cochannel interference or single-antenna intermodulation products fall near the receive channel frequency. Interference to FDM audio channels is avoided in these cases by setting a lower limit to the multiplex baseband and establishing appropriate transmitter and receiver local oscillator frequency tolerances. It is worth mentioning that similar consideration must be given to intermodulation products in the FDM baseband system. In the FDM system, all channel carrier and pilot frequencies are integer multiples of 4 kHz. This assures that all pilot intermodulation products will appear outside (0 or 4 kHz) the normal 0.3- to 3.4-kHz audio channel frequency limits.

As an example, consider a 6-GHz radio system with transmit and receive oscillator frequency accuracy of 0.001 percent. Assume a typical CCIR modulation section with six radio hops. A single-transmit desired carrier signal can be a maximum of 60 kHz ($6{,}000{,}000 \times 0.00001$) from nominal channel carrier frequency. Prior to demodulation at the end of the modulation section, it will be shifted randomly in frequency. If all frequency shifts are considered to be randomly distributed (Gaussian distribution with zero mean), the frequency shift at the FM demodulator due to six transmitters and six receiver oscillators will be $\sqrt{(6+6)}$ times the frequency shift of a single oscillator. The expected frequency error of the received carrier at the end of the modulation section is about 200 kHz ($60 \times \sqrt{12}$). If an interfering carrier (with similar frequency accuracy) appeared on the last hop, a tone in the demodulated baseband could be expected to appear between 200 and 260 kHz. Therefore, the FDM multiplex system should use a lower frequency limit of roughly 300 kHz.

Current high-density microwave radio transmission plans, which place all transmit frequencies at one end of the band and all receive frequencies at the other end, place some limitations on system design. When several transmit frequencies are used on a path, separate transmit and receive antennas are required to minimize interference due to waveguide intermodulation mixing products.

If a common antenna is used for both transmission and reception, careful planning of transmit and receive frequencies is necessary to avoid transmitter-to-receiver intermodulation crosstalk. Typically, separate antennas are required for more than three or four transmit frequencies on an antenna that is common with a receiver. Any time a common transmit and receive antenna is used with more than one transmitter, careful consideration must be given to transmitter intermodulation products that can appear in the receiver passband

The use of separate antennas takes advantage of the 60- to 110-dB coupling loss between side-to-side antennas pointed in the same direction. As with antenna front-to-back ratios, coupling loss can be seriously degraded by foreground reflections. Path terrain may require a zig-zag route due to reflections or over the horizon (overshoot) interference.

With modern frequency plans, a station at one end of a path is required to transmit using high frequencies and receive using low frequencies. The station at the other end of the path transmits low and receives high. If the paths form a closed loop, that loop must be composed of an even number of sites. If the closed loop is formed by an odd number of sites, one station will be required to transmit and receive both in the high- and low-frequency subbands. Such a location is called a bucking or bumping station. If a transmitter or receiver is operated at the same frequency at the same location, roughly 120-dB transmitter-to-receiver isolation is required to achieve acceptable levels of interference. This isolation can only be achieved at ideal locations with tall towers and no ground scatter or reflecting surfaces.

Transmitter-to-receiver signal leakage due to case, connector, or flange leaks is a common problem at these sites. Generally, this problem is solved by further subdividing the high and low subbands into individual transmit and receive subbands. This approach leaves unused channels for a guard band to reduce the adjacent channel interference. Since these approaches are not spectrally efficient, this path cannot carry as many channels as the other paths. Another approach is to place one of the loop paths in another frequency band. This approach is inefficient and is not always possible in highly developed areas. In general, the use of bucking stations should be avoided.

Frequency planning is started by the choice of a frequency plan. The use of plans must be carefully considered. The CCIR 1.7- to 1.9-GHz and 2.5- to 2.7-GHz frequency plans contain channel assignments that are outside the designated frequency band allocations. Several auxiliary channel assignments are outside the designated bands or are at or near main channel center frequencies. Due to designated frequency band limitations, CCIR channel bandwidths are not constant. In some cases, one or more channels may not be wide enough for the highest channel capacity applications without allowing some out-of-band emissions. Interleaved frequencies are spaced one-half channel from the main channel frequency. Sometimes the interleaved channels are one-half channel above and sometimes one-half channel below. Due to designated frequency band allocations, the number of interleaved channels is sometimes less than the number of main channels.

Detailed frequency planning is done on the basis of interference noise limits. The estimation of interference noise requires a knowledge of the desired signal (carrier) power, C, and the undesired interfering signal power, I. If the desired signal originates at station A, transmitting toward station B, and the interfering signal originates at station C, transmitting toward D, then the C/I observed at station B is given by

$$C/I(dB) = P(dB) + G(dB) + L(dB) + D(dB) \quad (4\text{-}1)$$

$$P(dB) = \text{transmitter power differential} \quad (4\text{-}2)$$

$$= P_c(dBm) - P_I(dBm) - L_c(dB) + L_I(dB) \quad (4\text{-}3)$$

$$G(dB) = \text{antenna gain differential} \quad (4\text{-}4)$$

$$= G_c(dB) - G_I(dB) \quad (4\text{-}5)$$

$$L(dB) = \text{free space loss differential} \quad (4\text{-}6)$$

$$= 20 \log (d_I/d_c) \quad (4\text{-}7)$$

$$D(dB) = \text{antenna discrimination} \quad (4\text{-}8)$$

$$= D_c(dB) + D_I(dB) \quad (4\text{-}9)$$

P_c	=	transmitter power of desired signal	(4-10)
P_I	=	transmitter power of undesired signal	(4-11)
L_c	=	power loss of desired signal between transmitter and transmit antenna	(4-12)
L_I	=	power loss of undesired signal between transmitter and transmit antenna	(4-13)
G_c	=	gain of transmit antenna at site A toward site B	(4-14)
G_I	=	gain of transmit antenna at site C toward site D	(4-15)
D_c	=	discrimination (relative to main lobe power) of receive antenna at site B toward site C	(4-16)
D_I	=	discrimination (relative to main lobe power) of transmit antenna at site C toward site B	(4-17)
d_c	=	distance from site A to site B	(4-18)
d_I	=	distance from site C to site B	(4-19)

For adjacent channel interference noise calculations on a multiline parallel route system, the C/I equation reduces to the combined cross-polarization discrimination (XPD) of the transmit and receive antennas. The combined XPD is never better than the worse of the two antennas. The data necessary to estimate terrestrial microwave interference is listed in Table 4-7.

Based on a calculated C/I, an estimate is made of interference noise. The interference depends on both the desired signal as well as the interfering signal. The preceding formula assumed free space transmission. Some interference cases may require calculation of obstruction or rain scatter loss. Adjacent channel interference requires an estimate of relative fading of the C and I signals. Occasionally, one or both signals will have a low frequency dispersal signal (burble) to obtain a burble factor interference reduction.

Table 4-7. Terrestrial frequency planning data.

1. Site name (with user identification).
2. Latitude: degrees, minutes, seconds, north or south.
3. Longitude: degrees, minutes, seconds, east or west.
4. Site elevation (meters or feet) above mean sea level.
5. Antenna center line (meters or feet) above site elevation - include data for both main and diversity antennas if appropriate.
6. Antenna description (manufacturer, type number (eg, UHX-10), type (eg, shrouded parabolic), feed type (eg, dual polarization horn), aperture diameter (eg, 10 feet) for main diversity antennas.
7. Antenna discrimination curves for both copolarization and orthogonal polarization (cross-polarized) signals.
8. Passive repeater size and type (eg, 10 feet by 10 feet, single billboard) and manufacturer and type number.
9. Equipment transmitter power and transmission line loss (or waveguide type and length) or transmitter power delivered to the transmit antenna.
10. Receiver transmission line loss (or waveguide type and length).
11. Transmitter frequency in MHz to nearest kHz (eg, 5945.200 MHz).
12. Transmitter frequency stability (eg, 0.005 percent).
13. Traffic type (video, telephony, data) and specific loading. If video, specify 525-line or 625-line, NTSC, SECAM, or PAL. If FDM telephony, indicate number of channels (eg, 960) and multiplexing plan (eg, CCITT Plan 2, 15 SGA). If data indicate bit rate and modulation method (eg, 90 Mb/s, 8 PSK) and if transmitter spectrum meets any emission mask.
14. Receiver interference susceptibility curves relating C/I to performance degradation for various cochannel and adjacent interfering signals.

FM Spectral Interference Noise - Theory

Excerpt from: Kizer, G. M., Microwave Communication, Ames: Iowa State University Press, 1990.

4.2 EXTERNAL INTERFERENCE NOISE

External interference in FM systems generally falls into one of several different categories. The first is interference caused by an interfering signal appearing at (cochannel interference) or very near (adjacent channel interference) the desired signal in such a manner that both appear at the FM demodulator. Medhurst [255] has developed a general formula for determining demodulated baseband signal distortion levels in FM systems experiencing this type of interference. This formula, applied to telephony systems, is as follows:

$$N(mW0) = \frac{1}{(C/I)\ (\Delta f_{pk})^2}\left\{\left[\frac{D^2}{p(D)}N_1\right] + \left[\frac{f^2}{p(f)}[N_2 + N_3 + N_4]\ df\right]\right\} \qquad (4\text{-}20)$$

$$= \frac{f^2}{2\ (C/I)\ (\Delta f_{/ch\ rms})^2\ p(f)}\ (N_1 + df\ [N_2 + N_3 + N_4]) \qquad (4\text{-}21)$$

since N_1 is much greater than $(N_2 + N_3 + N_4)$ when F equals D and N_1 is zero when f does not equal D.

$$N(dBm0) = 10 \log [N(mW0)] \qquad (4\text{-}22)$$

where

mW0 = power in milliwatts divided by the reference power (0 dBm0) in milliwatts (4-22.1)

$$N_1 = K_c K_I \delta(D - f) \qquad (4\text{-}23)$$

$$N_2 = K_c[M_I(|D - f|) + M_I(D + f)] \qquad (4\text{-}24)$$

$$N_3 = K_I[M_c(|D - f|) + M_c(D + f)] \qquad (4\text{-}25)$$

$$N_4 = \int_{-\infty}^{\infty}[M_c(|x + f|) + M_c(|x - f|)]\ M_I(|D - x|)\ dx \qquad (4\text{-}26)$$

All frequencies and powers must be measured in the same units (for example, kHz and mW0 per kHz).

$S_c(f)$ = spectrum power density of desired signal as it appears at the input of FM demodulator (ie, shaped by RF/IF section bandpass characteristics) (4-27)

$S_I(f)$ = spectrum power density of interfering signal as it appears at the input of FM demodulator (ie, also shaped by RF/IF section bandpass characteristics). S_c and S_I are both normalized to have unity total power and $f = f_{RF} - f_c$ (4-28)

f_{RF} = frequency at which the modulation spectrum power density is measured (4-29)

f_c = unmodulated carrier frequency of desired carrier (4-30)

f_I = unmodulated carrier frequency of undesired carrier (4-31)

f = baseband frequency of interest (center frequency of measurement) (4-32)

$\delta(x)$ = Dirac delta function (4-33)

= $(1, x = 0;\ 0,$ otherwise) (4-33.1)

D = $|f_I - f_c|$ (4-34)

= absolute difference in desired and undesired carrier frequencies (4-35)

C/I = desired carrier power/undesired carrier power [power ratio in N(mW0) in equation above, dB in N(dBm0) in equation which follows] (4-36)

Δf_{pk} = $\sqrt{2}\,\Delta f_{rms}$ (4-37)

Δf_{rms} = rms frequency deviation caused by reference (0 dBm0) sine-wave test-tone power level (4-38)

df = baseband noise power measurement bandwidth (narrow noise slot) (4-39)

N = unweighted noise relative to reference test-tone power measured in a noise slot with a width of df (4-40)

p(f) = baseband preemphasis power transfer characteristic (relative to pivot frequency) (4-41)

P(f) = 10 log [p(f)] (4-42)

$\Delta f_{/ch\ rms}$ = telephony per-channel rms deviation (for a reference 0 dBm0 sine-wave test tone) (4-43)

= Δf_{rms} (4-44)

$$S_c(f) = K_c\delta(f) + M_c(f) \tag{4-45}$$

$$S_I(f) = K_I\delta(f) + M_I(f) \tag{4-46}$$

where K_c and K_I are residual carrier power in the modulated spectrums and M_c and M_I are the modulated spectrum density functions as a function of frequency relative to unmodulated carrier frequency. If the carriers are both unmodulated, $K_c = K_I = 1$ and $M_c = M_I = 0$. These general formulas are valid for either FDM or digital interference spectrums.

N_1 represents a relatively strong spike of coherent noise power at the frequency $f = D$ in the receiver baseband. This spike (and its second and third harmonics) can cause considerable interference to normal signals as well as out-of-band noise and pilot sensors. N_2 represents modulation transferred from the interfering signal to the desired signal. This cross-modulation is caused by the presence of the residual carrier spike in the spectrum of the desired signal. N_3 represents modulation transferred from the desired signal to the undesired signal. Since the undesired signal's modulation spectrum is assumed to fall within the passband of the receiver, the transferred signal is then demodulated along with the desired modulation. This demodulated noise component would not exist if the undesired signal did not have a residual carrier spike. N_4 is similar to N_2 and N_3 in that the noise is generated by cross-modulation of the spectrums of desired and undesired signals. The desired and undesired modulated signal spectrums could be considered to be divided into frequency segments df wide. The continuous spectrums would be replaced by a spectrum of signal spikes df apart, each carrying the power in the df portion of the continuous spectrum. N_4 would be the limiting case (as df becomes 0) of the summation of all frequency cross-product components, which would fall within df/2 of the measurement frequency, f.

To consider the effect of FM filtering prior to demodulation, redefine S_c and S_I as spectrums as they appear at the input to a radio. The normalized power transfer function H(f) of the receiver between the RF signal input and the FM demodulator will be defined such that $H(0) = 1$ and $H(\infty) = 0$. For convenience, it will be assumed that the H(f) is symmetric about $f = 0$ [ie, $H(f) = H(-f)$]. For the high frequency predemodulation circuitry used in FM receivers, this is approximately true. As before, f is a predetection frequency relative to the unmodulated carrier frequency ($f = f_{rf} - f_c$) or a baseband frequency of interest (center frequency of a noise measurement using a measurement noise slot df wide).

$$N_1 = H(D)\, K_c\, K_I\, \delta(D - f) \tag{4-47}$$

$$N_2 = K_c[H(f)\, M_I\, (|D - f|) + H(f)\, M_I\, (D + f)] \tag{4-48}$$

$$= H(f)\, K_c\, [M_I(|D - f|) + M_I\, (D + f)] \tag{4-49}$$

$$N_3 = H(D)\quad K_I\, [H(|D - f|)\, M_c\, (|D - f|) + H(D + f)\, M_c\, (D + f)] \tag{4-50}$$

$$N_4 = \int_{-\infty}^{\infty}[H(|x + f|)\, M_c\, (|x + f|) + H\, (|x - f|)\quad M_c\, (|x - f|)]\; H(|x|)\, M_I\, (|D - x|)\, dx \tag{4-51}$$

If the spectrums are symmetric and bandlimited by frequency f_c such that

$$M_C(f) = M_C(-f) \text{ for all } f \tag{4-52}$$

$$M_I(f) = M_I(-f) \text{ for all } f \tag{4-53}$$

and

$$M_C(f) = M_I(f) = 0 \text{ for all } f \text{ such that } |f| \geq f_T \tag{4-54}$$

then

$$N_4 = \int_0^{(f_T + f)} \Big\{ H\,(x)\, H\,(f + x)\, M_c\,(f + x)\, M_I\,(|D - x|) + H\,(x)\, H\,(|f - x|)\, M_c\,(|f - x|)\, M_I\,(|D - x|) + H\,(x)\, H\,(|f - x|)\, M_c\,(|f - x|)\, M_I\,(D + x) + H\,(x)\, H\,(f + x)\, M_c\,(f + x)\, M_I\,(D + x) \Big\}\, dx \tag{4-55}$$

$$= \sum_{n=1}^{m} \Big[\Big\{ H\,(x)\, H\,(f + x)\, M_c\,(f + x)\, M_I\,(|D - x|) + H\,(x)\, H\,(|f - x|)\, M_c\,(|f - x|)\, M_I\,(|D - x|) + H\,(x)\, H\,(|f - x|)\, M_c\,(|f - x|)\, M_I\,(D + x) + H\,(x)\, H\,(f + x)\, M_c\,(f + x)\, M_I\,(D + x) \Big\}\, \Delta x \Big] \tag{4-56}$$

$$\text{for } \Delta x = [f_T + f]/m \tag{4-57}$$

$$x = [f_T + f]\,[n/m] \tag{4-58}$$

and m suitably large.

At this point, the expression for the FM system interference noise is quite general. The interfering signal can have any symmetric analog or digital spectrum S_I. Since the interference noise can be approximated by a summation, analytic expressions for the desired and interfering spectrums are not required. Photographs of spectrum analyzer displays can be used to obtain values for M_c and M_I. However, the displayed values must be normalized to the unmodulated carrier power. All discrete spectral components can be read as shown. All values of the continuous portions of the spectrums must be divided by the noise bandwidth of the spectrum analyzer.

If neither the desired carrier nor the interfering carrier is modulated,

$$M_2 = M_3 = M_4 = 0 \tag{4-59}$$

and

$$M_1 = 1 \text{ for } f = D \tag{4-60}$$

$$M_1 = 0 \text{ for } f \neq D \tag{4-61}$$

Therefore, the interference noise relative to the power of the reference baseband's sine wave (which causes the reference per-channel rms frequency deviation) is the following:

$$N(\text{dBm0}) = -3\ (\text{dB}) - C/I\ (\text{dB}) + 20 \log [D(\text{kHz})] - 20 \log [\Delta f_{/\text{ch rms}}\ (\text{kHz})] - P(D)(\text{dB}) + 10 \log [H(D)] \tag{4-62}$$

where C/I is the desired carrier-to-interfering carrier power ratio measured at the FM receiver input and H(D) is the overall receiver predetection rejection at the interfering carrier frequency relative to the receiver predetection response at the desired carrier frequency (10 log [H(D)] < 0). Fig. 4-1 plots this result for negligible IF filtering at the frequency of interest (the typical case) and no emphasis [P(D) = 0].

If either carrier is modulated, the situation becomes more complicated. To simplify matters, assume both the desired and interfering

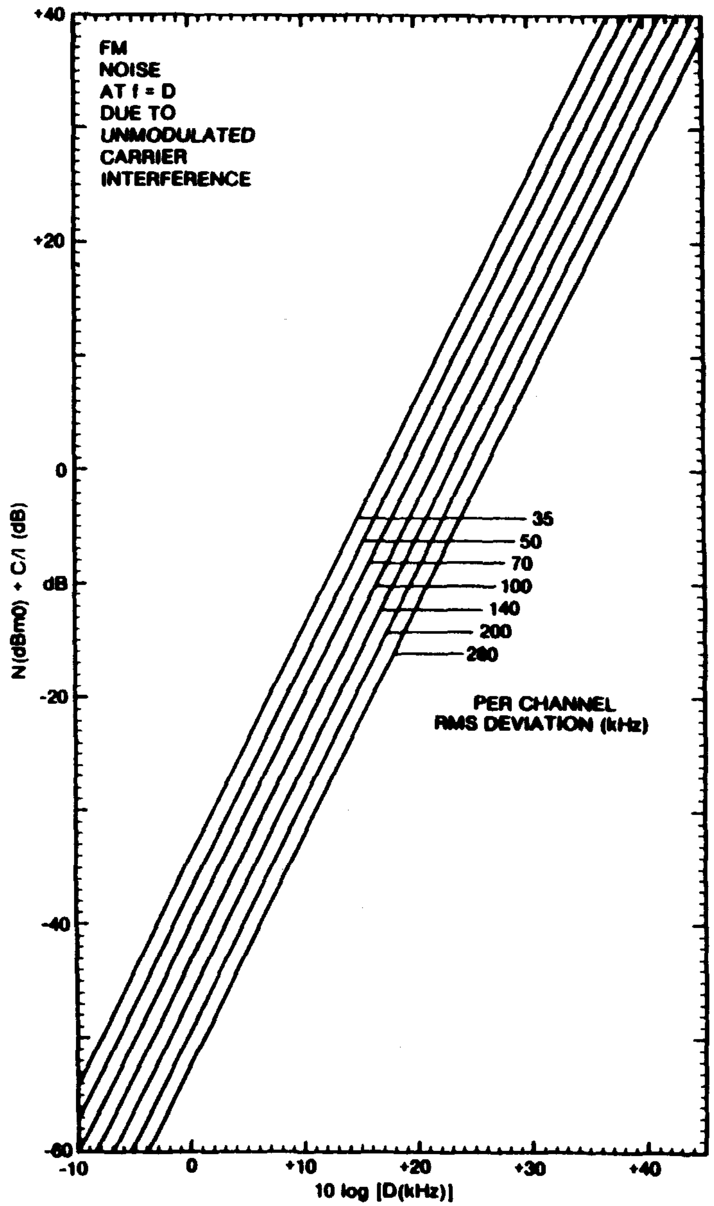

FIG. 4-1 Unmodulated carrier interference.

signals are identical types of FM signals. It is assumed that the basebands are described by Gaussian white noise signals bandlimited so the modulation signal extends from a low baseband frequency f_a to a high baseband frequency f_b. This, of course, is the typical representation of a frequency division multiplexed telephony baseband. It will be convenient to work with normalized spectral power density functions $G(f/f_b)$ such that

$$S\,(f/f_b) = \left[\frac{1}{f_b}\right] G\,(f/f_b) \tag{4-63}$$

If the desired carrier is modulated

$$S_c(f/f_b) = \left[\frac{1}{f_b}\right] G\,(f/f_b) \tag{4-64}$$

and if the interfering carrier is modulated

$$S_I(f/f_b) = \left[\frac{1}{f_b}\right] G\,(f/f_b) \tag{4-65}$$

The previous functions become the following

$$N_1 = H\,(D/f_b)\,K_c K_I\,\delta([D - f]/f_b) \tag{4-66}$$

$$N_2 = \frac{1}{f_b} N_6 \tag{4-67}$$

$$= \frac{1}{f_b} H\,(f/f_b)\,K_c\,[G_I\,(|D - f|/f_b) + G_I\,([D + f]/f_b)] \tag{4-68}$$

$$N_3 = \frac{1}{f_b} N_7 \tag{4-69}$$

$$\begin{aligned} = \frac{1}{f_b} H\,(D/f_b)\,K_I\,[&H\,(|D - f|/f_b)\,G_c\,(|D - f|/f_b) \\ &+ H\,([D + f]/f_b)\,G_c\,([D + f]/f_b)] \end{aligned} \tag{4-70}$$

$$N_4 = \frac{1}{f_b} I\,(f, G, H) \tag{4-71}$$

$$\begin{aligned} = \frac{1}{f_b}\int_{-\infty}^{\infty} [\,&H\,(|f + x|/f_b)\,G_c\,(|f + x|/f_b) \\ &+ H\,(|f - x|/f_b)\,G_c\,(|f - x|/f_b)] \\ &H\,(|x|/f_b)\,G_I\,(|D - x|/f_b)\,(dx/f_b) \end{aligned} \tag{4-72}$$

$$\begin{aligned} = \frac{1}{f_b}\int_0^{(f + f_T)/f_b} \Big\{ &H\,(\tau)\,H\,([f/f_b] + \tau)\,G_c\,([f/f_b] + \tau)\,G_I\,(|[D/f_b] - \tau|) \\ &+ H\,(\tau)\,H\,(|[f/f_b] - \tau|)\,G_c\,(|[f/f_b] - \tau|)\,G_I\,(|[D/f_b] - \tau|) \\ &+ H\,(\tau)\,H\,(|[f/f_b] - \tau|)\,G_c\,(|[f/f_b] - \tau|)\,G_I\,([D/f_b] + \tau) \\ &+ H\,(\tau)\,H\,([f/f_b] + \tau)\,G_c\,([f/f_b] + \tau)\,G_I\,([D/f_b] + \tau)\Big\}\,d\tau \end{aligned} \tag{4-73}$$

$$\begin{aligned} = \frac{1}{f_b}\sum_{n=1}^{m} \Big[\Big\{ &H\,(x)\,H\,([f/f_b] + x)\,G_c\,([f/f_b] + x)\,G_I\,(|[D/f_b] - x|) \\ &+ H\,(x)\,H\,(|[f/f_b] - x|)\,G_c\,(|[f/f_b] - x|)\,G_I\,(|[D/f_b] - x|) \\ &+ H\,(x)\,H\,(|[f/f_b] - x|)\,G_c\,(|[f/f_b] - x|)\,G_I\,([D/f_b] + x) \\ &+ H\,(x)\,H\,([f/f_b] + x)\,G_c\,([f/f_b] + x)\,G_I\,([D/f_b] + x)\Big\}\,\Delta x\Big] \end{aligned} \tag{4-74}$$

with

$$\Delta x = [f_T + f]/[m f_b] \tag{4-75}$$

$$x = n[f_T + f]/[m f_b] \tag{4-76}$$

m suitably large and

$$G_c\,(f/f_b) = G_I\,(f/f_b) = 0 \text{ for } |f/f_b| > f_c/f_b \tag{4-77}$$

$$N(\mathrm{dBm0}) = 10\log\left[\frac{(f/f_b)^2\,(N_1 + [df/f_b]\,[N_6 + N_7 + I])}{2(C/I)\,(\Delta f_{/ch\ rms}/f_b)^2\,p(f)}\right] \tag{4-78}$$

$$\begin{aligned} &= 10\log\,(N_1 + [df/f_b]\,[N_6 + N_7 + I]) \\ &+ 20\log\,(f/f_b) - 20\log\,(\Delta f_{/ch\ rms}/f_b) \\ &- 3 - C/I\,(\mathrm{dB}) - P\,(\mathrm{dB}) \end{aligned} \tag{4-79}$$

To generalize results, noise power ratio (NPR) will be used

$$NPR(dB) = NLR\ (dBm0) - BWR\ (dB) - N(dBm0) \quad (4\text{-}80)$$

$$BWR\ (dB) = 10 \log \left[\frac{f_b - f_a}{df}\right] \cong 10 \log \left[\frac{f_b}{df}\right] \quad (4\text{-}81)$$

for f_b much larger than f_a and df the (noise power) width of a telephone channel (nominally 3.1 kHz by convention)

NLR(dBm0) = 10 log [total baseband average power at the modulator input/reference sine-wave power] (4-82)

$$= 10 \log \left[\frac{[(\text{total rms frequency deviation})^2]}{[(\text{reference per channel rms frequency deviation})^2]}\right] \quad (4\text{-}83)$$

$$= 10 \log \left[\frac{\sigma_{rms}}{\Delta f_{/ch\ rms}}\right]^2 \quad (4\text{-}84)$$

$$= 20 \log \left[\frac{\sigma_{rms}}{\Delta f_{/ch\ rms}}\right] \quad (4\text{-}85)$$

Therefore,

$$NPR(dB) = 10 \log \left[\frac{2\ (C/I)\ (\sigma/f_b)^2\ p\ (f)}{(f/f_b)^2\ [N_5 + N_6 + N_7 + 1]}\right] \quad (4\text{-}86)$$

$$= 3 + C/I\ (dB) + P\ (dB) + 20 \log (\sigma/f_b) - 20 \log (f/f_b) \quad (4\text{-}87)$$

$$- 10 \log [N_5 + N_6 + N_7 + 1]$$

f = baseband frequency of interest (4-88)

f_b = highest baseband frequency (4-89)

= frequency of highest FDM telephone channel (4-90)

df = telephone channel (noise) bandwidth (4-91)

≅ 3.1 kHz by convention (4-92)

$\Delta f_{/ch\ rms}$ = FM rms frequency deviation caused by a reference 0 dBm0 sine-wave (4-93)

C/I (dB) = desired carrier to interfering carrier power (expressed in dB) at the input to the FM receiver (4-94)

P (f) (dB) = preemphasis power transfer characteristic (expressed in dB) relative to pivot frequency (4-95)

$$20 \log (\sigma/f_b) = 10 \log (\sigma/f_b)^2 \quad (4\text{-}96)$$

$$= 10 \log (\sigma/\Delta f_{/ch\ rms})^2\ (\Delta f_{/ch\ rms}/f_b)^2 \quad (4\text{-}97)$$

$$= 10 \log (\sigma/\Delta f_{/ch\ rms})^2 + 20 \log (\Delta f_{/ch\ rms}/f_b) \quad (4\text{-}98)$$

$$= NLR\ (dBm0) + 20 \log (\Delta f_{/ch\ rms}/f_b) \quad (4\text{-}99)$$

NLR (dBm0) = power of bandlimited white noise baseband loading relative to reference sine-wave power (4-100)

= $-4 + 6 \log N$ for $N < 240$ (4-101)

= $-15 + 10 \log N$ for $N \geq 240$ per CCIR/CCITT recommendations (4-102)

$$N_5 = (f_b/df)\ N_1 \quad (4\text{-}103)$$

$$= 0 \text{ when } f \neq D \quad (4\text{-}104)$$

$$= \text{dominant noise when } f = D \quad (4\text{-}105)$$

The previous formulas have been derived assuming FM. What started as baseband powers into an unemphasized modulator and out of an unemphasized demodulator were converted into rms frequency deviations σ and $\Delta f_{/ch\ rms}$. Although the following sections will consider preemphasized FM as well as practical PM, the values of σ and $\Delta f_{/ch\ rms}$ will not be changed. At first this might seem strange since using emphasis or baseband frequency response shaping network between bandlimited white noise and the modulator input will certainly

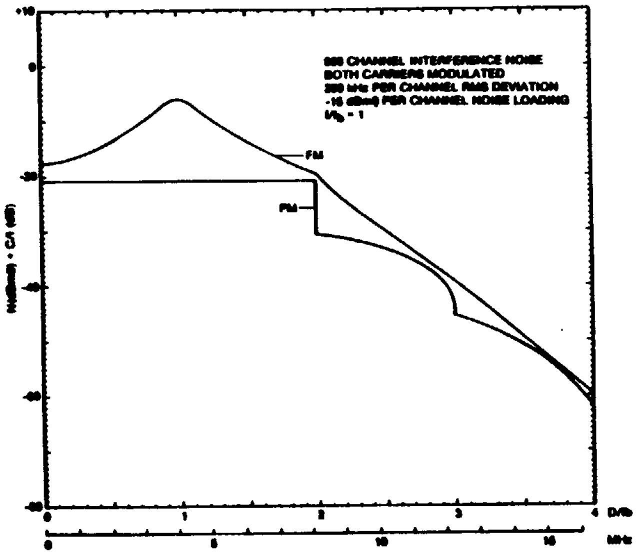

FIG. 4-21 960-channel interference noise.

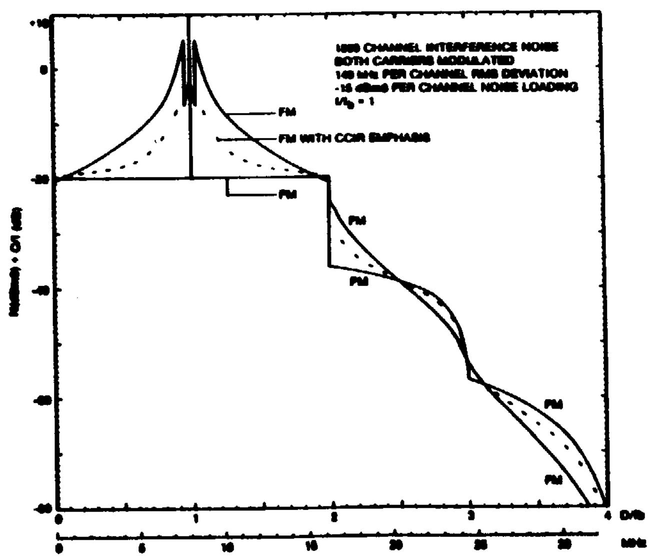

FIG. 4-22 1800-channel interference noise.

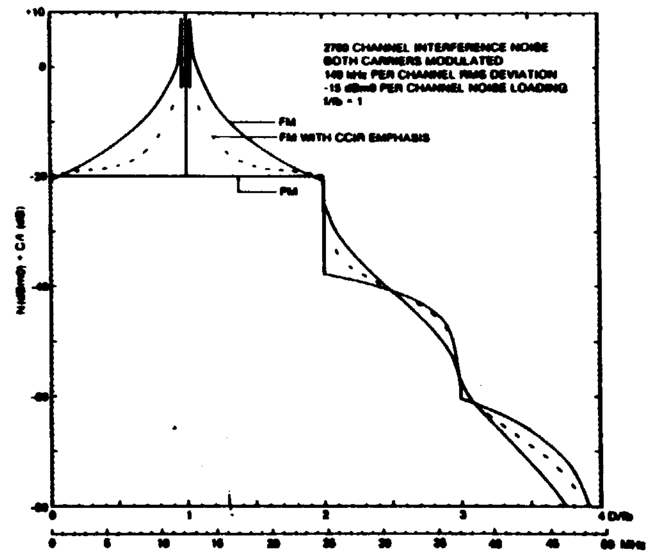

FIG. 4-23 2700-channel interference noise.

These results are the same as the previous $f_a = 0$ results for $|D| \leq 2\,f_b$. For $2\,f_b < |D| \leq 3\,f_b$, these results predict 2.5 dB lower noise and for $3\,f_b < |D| \leq 4\,f_b$, these results predict 5 dB lower noise than for the previous $f_a = 0$ results. This difference is due to the dependency of the higher order FM spectrum on f_a/f_b for the low deviation ($\sigma/f_b < 1/4$) case.

D will vary slightly from path to path due to slight tuning inaccuracy even though D has the same nominal value for each path. Therefore, adjacent channel interference noise from one path to the next will be uncorrelated. This noise will tend to build up over multiple hops by 10 log n where n is the number of radio hops (paths) in a row. This type of interference occurs for baseband or IF/RF repeaters.

Most FM receivers use a single heterodyne or mixing process to transfer the desired signal at frequency f_c to a more convenient intermediate frequency f_{IF} for further processing. Such a receiver is prone to interference from signals at frequencies far removed from the desired frequency f_c [157], [158], [186], [420]. The receiver is most susceptible to three types of spurious response due to the nonlinear mixing process. The first type is multiple image interference. If f_{LO} is the frequency of the receiver local oscillator used in the mixing process and n is an integer, the receiver will produce an f_{IF} signal for an

FM Spectral Interference Noise - Practice

Technical Considerations for Digital Expansion of Analog FM 2400 Channel Multiline 6 GHz Microwave Systems

George Kizer

INTRODUCTION

Over several years several of the North American telephony common carrier microwave transmission systems evolved to a 2400 channel L4 carrier FDM multiline 6 GHz FM analog transmission system. The rapid expansion of the system to include digital transmission has caused the need for expansion of existing analog routes using digital transmission facilities. This paper describes the technical considerations for implementing such an expansion on an existing high capacity analog system.

FM INTERFERENCE CASES

Multiline radio transmission systems are subject to several forms of radio signal interference. System engineering and frequency planning is using to minimize interference due to converging RF routes, spurious receiver responses, and waveguide nonlinearity mixing. Multiline FM systems are subject to three forms of interference which are a function of FM transmission parameters and RF channel spacing. This interference must be considered prior to implementation since little can be done after the fact to reduce it. Theoretical examples will be given for typical digital transmission systems. The results on an actual implementation of a 1000 mile route system will be mentioned.

CASE 1: DIRECT ADJACENT CHANNEL INTERFERENCE NOISE

Interference from overlapping sidebands of interfering signals will be garbled or unintelligible since the disturbing sidebands are not correlated with the disturbed carrier. In systems with closely spaced FM channels, however, a more complicated form of adjacent channel interference has been noted in which the interference is intelligible. This type of interference, in which the signal on the adjacent channel appears as an identical signal in the disturbed channel, has been termed Direct Adjacent Channel Interference (DACI). The transfer mechanism has been described by Ruthroff [18], Chapman and Millard [3], and Goldman [6]. Curtis, Collins, and Jamison [4] observed that the interference appears in the disturbed channel at exactly the same baseband frequency as applied to the other channel. A change of 1 dB in the carrier ratio of the two channels produces a 2 dB change in the baseband interference in the disturbed channel provided the disturbing carrier is weaker than the disturbed carrier. A change of 1 dB in the level of the baseband input to the disturbing channel produces a change of 1 dB in the interference observed in the disturbed output. The baseband interference was observed to be essentially independent of the modulation frequency on the disturbing carrier. With a fixed difference in level between the disturbing and disturbed carriers, the interference as observed at baseband is independent of the power of the disturbing carrier at the input of the radio receiver. Experimental work indicates that the crosstalk mechanism which is characteristic of adjacent channel interference exists in the limiter of the FM receiver and also at any point in the system where there is a pronounced tendency to compress the signal. The amplitude of this interference varies as the square of the ratio of the voltage at which the limiter clips to the signal voltage impressed upon it. The most effective method of reducing this type of interference appears to be filtering out the disturbing signal before it reaches a limiter or compressing amplifier.

Since this adjacent channel noise mechanism has been identified in FM systems, it should be considered in applications of digital adjacent channel applications. Experience with analog FM adjacent channel systems indicates that if adjacent channel filtering and/or radio frequency (RF) cross polarization discrimination is adequate to control Tertiary

Interference, DACI due to an FM system will be insignificant. If the adjacent channel is digital, the DACI mechanism will not produce intelligible noise and its effect will be less than that of an analog system. Noise due to this mechanism is therefore considered negligible for digital adjacent channel operation.

CASE 2: TERTIARY INTERFERENCE

Another type of interference, called Limiter Transfer or Teriary Interference, is most likely to occur when a multiple RF channel system is implemented along a common path. The sidebands of one channel may be picked up by a second carrier (transfer) channel and then transferred to a third channel through limiter action. This causes signals from the first channel to appear in the third channel without appearing in the second channel. Limiter Transfer may be explained by considering the effect of a limiter on an input signal composed of two unmodulated carriers [5], [8], [10], [11], [13], [19]. The larger of the two will be called the desired carrier and the smaller the interference. When these signals are passed through a limiter, both signals are reduced in amplitude and many other signals are generated. The limiter output consists of signals in the same frequency range as the input signals as well as all odd (eg, third, fifth, seventh) harmonics of the input signals. Also, the output consists of signals at the carrier frequency plus and minus all integral multiples of the frequency difference between the two input frequencies. In general, the odd-order harmonics are filtered out and the sum and difference frequencies greater than ±1 are low enough to be ignored. The dominant output of the limiter, therefore, is a main carrier plus two equal sidebands, one at the original interfering signal frequency and another on the other side of the carrier. Each will have the same absolute frequency difference between it and the main carrier but each will be 6 dB lower in amplitude relative to the carrier than was the original interfering signal.

In a parallel route multiple RF channel system (see Figure 1), an interfering signal may appear on one side of a carrier, the composite signal limited, and then the resultant signal retransmitted. Upon retransmission, the interference appears on both sides of the transfer carrier. Although the prelimiting interference may be too far removed from the desired carrier to produce significant distortion in the carrier's receiver, the new interfering signal may cause considerable interference to the RF channels next to it. This can occur even though the RF channel being interfered with has adequate filtering and appropriate frequency planning to avoid untransferred interfering signals. In long parallel route systems, it is possible for the interference to be transferred onto other channels as well as back into the original channel (on another RF hop). For long parallel routes with equally spaced RF channels, this type of interference can build up and cause erratic behavior of high frequency pilot and switch control tones. Multiline switch problems can occur if noise sensor slots are so close to pilot frequencies that pilots being transferred back into the originating route (with a slight frequency effort due to slight mistuning of oscillator frequency) fall into the noise slot. As with DACI, the most effective way to avoid this type of interference is to filter out the potentially interfering signals before they are applied to a limiter of the potential carrier (transfer) channel. This type of interference can occur for remodulation of RF/IF repeater systems.

Jansen and Prime [9] have described the mechanism for Tertiary Interference in an RF channel of a parallel route multiple RF channel telephony FM system. Two general types of interference occur. Consider the three RF channels shown in Figure 1 (actual parallel route systems generally have 6, 8, or 12 RF channels when fully developed). The first-order FM sidebands of channel 1 (from site A), after having passed across path 1, appear at the input to the limiter of channel 2 at site B. The level of the sidebands will be reduced by the composite filtering characteristic of the repeater and cross-polarization discrimination of the receive antenna. After passing through the limiter, the sideband energy is reduced 6 dB but transferred to both sides of the channel 2 signal transmitted toward site C. The interfering first-order sideband energy that appears back in channel 1 is called double adjacent tertiary interference. The interfering energy that appears in channel 3 is called transfer tertiary interferences. If the interference occurs on two adjacent paths (eg, from channel 1 to channel 2 on path 1 and from channel 2 to channel 1 or 3 on path 2), the interference is called first order. If the interfering sidebands pass through two repeaters rather than one before appearing in the other channel, the interference is called second order. For a system of N hops, up to (N-1) orders of interference are possible. If the assumption is made that the signal-to-interference ratio in the first-order sideband region after limiting is equal to the NPR at baseband, the noise due to all orders of tertiary interferences would be given by

$$NPR_T\,(dB) = 12.0 + 2\,C/I\,(dB) - T - 10 \log n$$

where

$$T = 10 \log \sum_{K=1}^{N-1} (N-K)\, a^K$$

C/I(dB) = carrier-to-interference ratio of a path prior to filtering and limiting

a(pr) = loss (expressed as a power ratio) of a repeater filter at frequency (D - f) away from the repeater unmodulated carrier frequency relative to the filter loss at the unmodulated carrier frequency (note that $0 \leq a \leq 1$)

D = unmodulated carrier frequency difference between any two adjacent RF channels on the same path

f = baseband frequency of interest

L(dB) = 10 log (a)

N = number of parallel route hops

n = number of interference paths (note that n is 1, 2, or 4)

Figure 2 plots the combined IF and RF power response of a typical 2400 channel FM receiver. The dominant intermodulation distortion contribution due to even order differential gain or phase (eg, bow) occurs over the first and second order FM sidebands (within plus or minus 2 f_b of the carrier frequency) [17]. Therefore, the FM receiver response of Figure 2 represent the practical limit of FM filtering prior to detection.

If a parallel route system consisted of only two parallel routes, n would be 1 because each channel would interfere with itself through double adjacent tertiary interference. Transfer interference would not be possible for this case. If a parallel route system consisted of only three parallel routes (see Figure 1), n would be 2 for all three channels. Channel 1 would receive transfer tertiary interference from channel 3 and double adjacent tertiary interference from itself. Channel 3 would receive tertiary interference from channel 1 and double adjacent interference from itself. Channel 2 would not receive any transfer tertiary interference but would receive double adjacent channel interference due to both channels 1 and 3. In the general parallel route case, n is 4 because a channel receives both transfer and double adjacent interference from each side of the path. Note that this type of interference only occurs in systems with IF or RF repeaters. The process terminates every time the signal is demodulated to baseband. The interference process starts again on the next baseband-to-baseband section of multiple IF or RF repeaters. The noise on n identical baseband switch sections would add on a 10 log n basis.

The preceding analysis assumed a homogeneous analog FM multiline system. This analysis can be extended to a mixed digital and analog situation [12]. In the preceding example, if channel 1 is digital and 2 and 3 are analog FM channels, channel 1 will produce transfer tertiary interference into channel 3. The analysis would be the same as for an analog system except for the addition of a factor U to account for the relation between the modulated spectrum sideband energy of the FM and digital signals. It is assumed that the digital signal is known to produce S power relative to total carrier power when measured in a noise slot df wide located f away from carrier frequency. Figure 4 plots values of S for df of 4 kHz for various practical telephony digital spectrums. For comparison, Figure 3 plots the same for angle modulated analog transmission [7], [12], [15], [16]. Figure 3 results were taken from [12] results by multiplying the normalized spectrums by 10 $\log(df/f_b)$. Results have been given for unemphasised FM, FM with CCIR emphasis, and for PM. Results have been given for PM since this represents the optimum angle modulation system when evaluated using an external interference rejection criterion. Figure 10 shows the emphasis characteristics of CCIR and FM to PM conversion emphasis curves. The PM results may be considered the theoretical best results that can be obtained from an angle modulation system.

If only the highest baseband frequency f_b is considered, the transfer tertiary interference due to a digital system is give by [12]

$$NPR_T\,(dB) = 12.0 + 2\,C/I\,(dB) - T - 10 \log n - U$$

where for FM without emphasis

$$U = 10 \log (S) + 3 + 10 \log (f_b/df) - 20 \log (\sigma/f_b)$$

or for FM with CCIR emphasis

$$U = 10 \log (S) - 1.8 + 10 \log (f_b/df) - 20 \log (\sigma/f_b)$$

Consider an example of a 2400 channel FDM-FM system with typical North American 6 GHz channel spacing. This RF channelization conforms to Part 21.701 of the Code of (United States) Federal Regulations, Title 47, Chapter I and CCIR Recommendation 383-3. This frequency plan assigns eight go and eight return channels with 29.65 MHz channel bandwidth. For this channelization NPR_T for an all analog 2400 channel system conforming to the characteristics shown in the Appendix (N=3) is the following

$$NPR_T\,(FM\text{-}FM) = 12 + 2\,C/I - 10 \log (3) - 10 \log (4)$$

$$= 1.2 + 2\,C/I(dB)$$

The switching section (3 hop) tertiary interference objective of +8.0 dBrnCO is equivalent to an unweighted signal to noise ratio (S/N) of +80.0 dB. The equivalent NPR is given by [12]

$$NPR(dB) = S/N(dB) - 16.0 + 10 \log (2400) - 10 \log ((11404-564)/3.1)$$

$$= 62.4$$

To achieve the noise objective, the per hop C/I objective must be 31 dB.

Adjacent channels of a multiline microwave radio system are normally cross polarized. The received C/I is the combined cross polarization discrimination of the transmit and receive antennas. If all antennas are assumed to be the same, C/I is twice the cross polarization discrimination of a single antenna. Therefore, this objective can be met using microwave antennas with 16 dB cross polarization discrimination.

For digital interference, using data from Figure 4 and D-f = 28.65 - 11.40, the following U factors are obtained for emphasized systems.

$$U(8\ PSK) = -62.0 - 1.8 + 10 \log (11404/4) - 20 \log (0.0681)$$

$$= -17.6$$

$$U(16\ QAM) = -82.0 - 1.8 + 10 \log (11404/4) - 20 \log (0.0681)$$

$$= -37.6$$

$$U(64\ QAM) = -80.5 - 1.8 + 10 \log (11404/4) - 20 \log (0.0681)$$

$$= -36.1$$

For the above, the IF/RF receiver response at D-f was assumed to be 0 dB(per Figure 2), per channel rms deviation was 100 kHz and per channel white noise loading was -16 dBm0.

Clearly if an existing 2400 channel system meets tertiary interference objectives, tertiary interference due to an adjacent channel digital system will be negligible.

CASE 3: ADJACENT CHANNEL INTERFERENCE

Interference caused by an interfering signal appearing in frequency near the desired received FM signal is known as Adjacent Channel Interference. Medhurst [14] has developed a general formula for determining demodulated baseband signal distortion levels in FM systems experiencing this type of interference. This formula, applied to telephony systems [12] is as follows:

$$N(mw0) = f^2 [N_1 + df (N_2 + N_3 + N_4)] / [2 (C/I) (\Delta f/ch\ rms)^2 p(f)]$$

where

C/I = desired carrier power/undesired carrier power ratio

df = baseband noise power measurement bandwidth (narrow noise slot)

Δf/ch rms = telephony per-channel rms deviation (for a reference 0 dBm0 sine-wave test tone)

f = baseband frequency of interest (center frequency of measurement)

N = unweighted noise relative to reference test-tone power measured in a noise slot with a width of df

p(f) = baseband preemphasis power transfer characteristic (relative to pivot frequency)

N_1 represents a relatively strong spike of coherent noise power at the frequency f = D in the receiver baseband. This spike (and its second and third harmonics) can cause considerable interference to normal signals as well as out-of-band noise and pilot sensors. N_2 represents modulation transferred from the interfering signal to the desired signal. This cross-modulation is caused by the presence of the residual carrier spike in the spectrum of the desired signal. N_3 represents modulation transferred from the desired signal to the undesired signal. Since the undesired signal's modulation spectrum is assumed to fall within the passband of the receiver, the transferred signal is then demodulated along with the desired modulation. This demodulated noise component would not exist if the undesired signal did not have a residual carrier spike. N_4 is similar to N_2 and N_3 in that the noise is generated by cross-modulation of the spectrums of desired and undesired signals. The desired and undesired modulated signal spectrums could be considered to be divided into frequency segments df wide. The continuous spectrums would be replaced by a spectrum of signal spikes df apart, each carrying the power in the df portion of the continuous spectrum. N_4 would be the limiting case (as df becomes 0) of the summation of all frequency cross-product components, which would fall within df/2 of the measurement frequency, f.

Consider an example of a 2400 channel FDM-FM system similar to the one considered in the previous section. Adjacent channel interference noise has been calculated numerically for the highest baseband frequency (worst case) using receiver filtering shown in Figure 2. The method used is described in [12]. The digital spectrums S(f) of modulated power in a 4 kHz wide slot relative to the unmodulated carrier power was multiplied by 10 $\log[(f_b/df)(df/4kHz)]$ to generate the normalized interference spectrum $G_I(f)$. Results have been given for unemphasised FM, FM with CCIR emphasis, and for PM. Results have been given for PM since this represents the optimum angle modulation system when evaluated using an external interference rejection criterion. Figure 10 shows the emphasis characteristics of CCIR and FM to PM conversion emphasis curves. The PM results may be considered the theoretical best results that can be obtained from an angle modulation system. The numerical results are displayed in Figures 5, 6, 7, and 8. For D = 29.65 MHz and CCIR emphasis, the following unweighted signal to noise (S/N) values were obtained:

S/N (FM-FM) = 47.3 + C/I(dB)

S/N (FM-8 PSK) = 27.8 + C/I(dB)

S/N (FM-16 QAM) = 30.5 + C/I(dB)

S/N (FM-64 QAM) = 31.1 + C/I(dB)

Since tertiary interference is not a factor for adjacent channel digital signals, assume that the 80 dB unweighted signal to noise (S/N) objective (+8.0 dBrnC0) tertiary interference 3 hop objective may be used for the digital adjacent channel interference objective. This equates to a per hop S/N objective of 85.2 dB.

As noted previously, adjacent channels of a multiline microwave radio system are normally crosspolarized. The received C/I is the combined cross polarization discrimination of the transmit and receive antennas. If all antennas are assumed to be the same, C/I is twice the cross polarization discrimination of a single antenna. Therefore, the following antenna cross polarization (XPD) objectives must be met for the following types of adjacent signals.

XPD (FM-FM) = 19 dB

XPD (FM-8PSK) = 29 dB

XPD (FM-16QAM) = 27 dB

XPD (FM-64QAM) = 27 dB

These values are achievable but require XPD to be 8 to 10 dB better for digital adjacent channel operation.

ACTUAL EXAMPLE

In 1986 Rockwell International contracted with a major interexchange carrier to expand an existing 1000 mile 4 channel multiline 2400 channel FDM-FM system to 8 channels using 3 DS-3 64 QAM digital transmission. The system had 43 serial hops averaging 23.4 miles each. The system had 8 switching sections with 6 the median number of switching section hops. This would increase tertiary noise 6 dB relative to the previous calculations. This noise, however, is still insignificant. To confirm the theoretical calculations, measurements were made using the actual radios to be deployed onto the system. Those results are displayed in Figure 9. The actual results were in reasonable agreement with the theoretical results. In actual implementation, the entire system was optimized for the new digital channels. All antennas had cross polarization reoptimized. An objective of 35 dB was used for all orientations. When this could note achieved for both polarizations, the polarizations for digital into analog interference were optimized. After converting the system to mixed digital and analog operation, noise in all analog radio basebands was measured with and without digital signals present. Without the digital signals, the end to end noise objectives were met. No increase in baseband noise occurred at any frequency in the baseband when the adjacent channel digital signals were added. The noise performance of the system was, and continues to be, an unqualified success.

CONCLUSION

This paper evaluated the three dominant forms of interference to analog FM multiline microwave systems subjected to the spectrum of a typical adjacent channel digital signal. Adjacent Channel Interference due to spectrum overlap was found to be the dominant source of noise. Numerical calculations concluded that noise objectives could be met by increasing antenna crosspolarization discrimination (XPD) 8 to 10 dB relative to FM noise requirements. This theoretical result was actually tried in the field on a continuous 1000 mile multiline system. Outstanding results were obtained using an antenna XPD objective of 35 dB. Expansion of 2400 channel analog FM systems has been shown to be achievable both theoretically and in practice.

APPENDIX: ANALOG TRANSMISSION PERFORMANCE OBJECTIVES:

The following long-haul reference circuit information is based on Jansen and Prime [9] unless otherwise noted. The 4000-mile ATT hypothetical reference circuit is composed of 150 equal-distance radio hops. The system consists of 51 main stations with IF switching and 100 IF repeater stations (50 cascaded IF switching sections with two IF repeaters in each section). Of the 51 main terminals, 17 are multiplex terminals. These are assumed to be interconnected by FM terminals (FM modulator and demodulator pair), wireline entrance links, and multiplex (at channel, group, or supergroup level). The multiline switching section is 1 x N [20].

Worst-case end-to-end circuit noise for the 4000-mile system during periods of nonfaded transmission is 41 dBrnC0. The noise may increase to 55 dBrnC0 during fading, after which the channel will be switched automatically to a protection channel. Single-tone interference of -68 dBm0 (41 dBrnC0) is the maximum for any voice circuit in the 4000-mile circuit during unfaded transmission. Under normal conditions, this corresponds to -87 dBm0 per hop (only one-half the hops accumulate the tone). Subjective tests have shown that, if the noise-to-tone power ratio in a message circuit is constant, the tone is less discernible when the noise power is increased. The result is that the requirement for those baseband tones that increase dB for dB with fading is -47 dBm0 when the circuit noise is 55 dBrnC0 (during a 40-dB fade).

Unfaded end-to-end noise of 41 dBrnC0 (11,220 pW0p) has been allocated as follows:

a. +31.2 dBrnC0 for 16 pairs of channel, group, and supergroup multiplex. If frogging reduces intermodulation buildup to 10 log n, this equates to a single-pair multiplex objective of 19.2 dBrnC0 (74.1 pW0p).
b. +28.0 dBrnC0 for 16 pairs of wireline entrance links. If intermodulation buildup is 10 log n, this equates to a single-pair objective of 16 dBrnC0 (35.5 pW0p).
c. +29.0 dBrnC0 for 16 pairs of FM terminals (modulator and demodulator pair). If intermodulation buildup is 10 log n, this equates to a single-pair objective of 17 dBrnC0 (44.7 pW0p).
d. +39.9 dBrnC0 (8710 pW0p) for 150 hops of radio. The radio noise is subdivided as follows:
    1. +28 dBrnC0 for (same route) cochannel interference. This equates to +6.2 dBrnC0 (3.7 pW0p) per hop.
    2. +26 dBrnC0 for cochannel (converging route) interference. This equates to +4.2 dBrnC0 (2.3 pW0p) per hop.
    3. +29 dBrnC0 for intersystem interference. If two exposures per hop are assumed, this equates to +4.2 dBrnC0 (2.3 pW0p) per exposure per hop.
e. +28 dBrnC0 for RF echo distortion. This equates to +6.2 dBrnC0 (3.7 pW0p) per hop.
f. +22 dBrnC0 for IF (interconnect) echo distortion. With 100 IF interconnects, this equates to +2.0 dBrnC0 (1.4 pW0p) per IF hop.
g. +25 dBrnC0 for tertiary interference. For 50 IF switch sections, this equates to +8.0 dBrnC0 (5.6 pW0p) per 3-hop IF section.
h. +38.4 dBrnC0 to 150 microwave transmitter and receiver pairs. This noise is further subdivided as follows:
    1. +36.9 dBrnC0 for thermal noise. This equates to +15.1 dBrnC0 (28.8 pW0p) per hop (10 log n addition).
    2. +33.1 dBrnC0 for intermodulation noise. This equates to +2.6 dBrnC0 (1.6 pW0p) per hop. This is based on measured intrasection intermodulation addition of 19 log n yielding an end-to-end intermodulation noise buildup of 14 log n. It is interesting to note that intermodulation noise of cascaded identical devices has